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Abstract

This paper develops a technique for bounding the maximum voltages and currents at terminations of a multiconductor transmission line (MTL) located behind an aperture-perforated conducting screen excited by an electromagnetic field in the frequency domain. The electromagnetic field is coupled through a small aperture as the excitation of a multiconductor transmission line behind the aperture. A model is presented in terms of external and internal sources which in turn create traveling waves on the multiconductor transmission line. These traveling waves transfer energy to the terminations. The energy at a termination is translated into voltages and currents from which the upper bounds are determined. These upper bounds are obtained using vector norms and associated matrix norms. The formulation is presented in the frequency domain to obtain useful upper bounds for analysis of multiconductor transmission line geometries with aperture excitation.

Introduction

In designing some systems, the designer should be able to characterize the penetration of electromagnetic pulses (EMP) or lightning signals through apertures of general shapes as well as quantify the effects of the coupled energy on transmission lines located in the vicinity of the aperture. Apertures that are of concern to the designer are usually electromagnetically small over the spectrum of the EMP, or lightning, and their existence may be for some purpose, e.g., windows, open access holes, or they may be unintentional as in the case of cracks around doors or plates covering access ports or poor electrical seams. The analysis of the coupling (penetration) problem has been investigated by a large number of people since 1897. The first scientist to propose a solution was Lord Rayleigh [1], whose solution was expressed as an ascending power series of the wavenumber $k(= \frac{2\Pi}{\lambda})$ where λ is the wavelength. Others include Bethe [2], Bouwkamp [3], and more recently Butler [4].

The coupling of the energy from an incident electromagnetic wave to a transmission line located behind an aperture-perforated conducting screen has also been investigated by many engineers and physicists in the past decade. Kajfez [5] has computed the coupled energy by the use of equivalent electric and magnetic dipole moments along with both mode-matching and reciprocity techniques to obtain equivalent sources of a transmission line model. Butler and Umashankar [6] have approached the problem numerically by the method of moments, and have formulated integro-differential equations for a finite-length wire with arbitrary orientation behind an arbitrarily shaped aperture. Davis [7] has developed a method for bounding the maximum voltage and current levels at terminations of a single wire behind an aperture.

This paper extends the bounding problem of a single wire to the problem of obtaining an upper bound for the computation of the voltages and currents at terminations of multiconductor transmission lines (MTL) located behind an aperture-perforated conducting screen. A brief summary of the system model will be given, followed by a more detailed discussion of the signal computations and the resultant upper bounds.

System Model

A typical multiconductor transmission line (MTL) problem of interest is shown in Fig. 1. The aperture A is the source of coupled energy which is modeled by a pair of electric and magnetic current dipoles above a closed aperture as shown in Fig. 2. It is easily shown, from the small aperture theory as used by Kajfez [8], that these point source amplitudes are given by

$$\bar{c}_e = c_{ey}\hat{y} = j\omega\epsilon\alpha_e E_y^{sc-}\hat{y} \tag{1}$$

and

$$\bar{c}_m = c_{mx}\hat{x} = -j\omega\mu\alpha_{m,xx} H_x^{sc-}\hat{x} \tag{2}$$

respectively, where ω is radian frequency, α_e and $\alpha_{m,xx}$ are the required polarizabilities of the aperture and $\bar{E}^{sc-}$ and $\bar{H}^{sc-}$ are the electromagnetic fields below the aperture with no aperture present. The H_z^{sc-} is of no importance to the problem and has been neglected, as have the low level aperture fields of the MTL. In the process of bounding, E_y^{sc-} and H_x^{sc-} are assumed to be given while the α_e and $\alpha_{m,xx}$ may either be given by the appropriate geometry related coefficients or bounded by the polarizabilities of an ellipse which circumscribes the aperture [7].

We will restrict the MTL to be in a vacuum environment and to consist of N parallel wires. In such a case there are N transverse

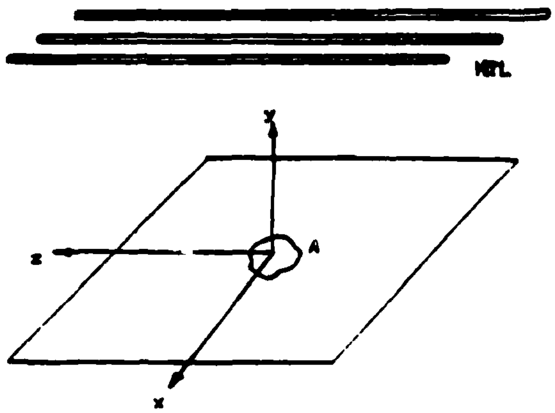

Figure 1. A multiwire transmission line (MTL) parallel to a plane with aperture A.

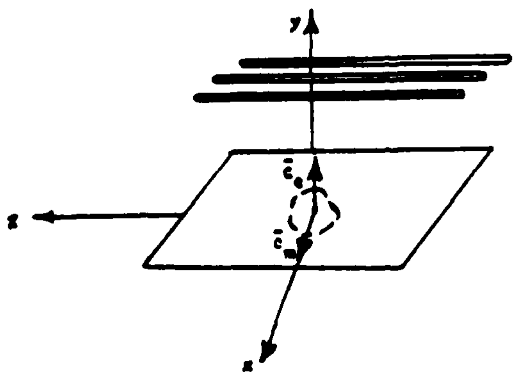

Figure 2. Model of the aperture as replaced by equivalent current dipoles.

electromagnetic (TEM) modes possible on the MTL. The equations that govern these modes are

$$\frac{d}{dz} |V> = -j\omega \underline{L} |I> \tag{3}$$

and

$$\frac{d}{dz} |I> = -j\omega \underline{K} |V> \tag{4}$$

where $|V>$ and $|I>$ are vectors representing the N wire voltages and N z-directed wire currents. The induction matrix $\underline{L}$ and capacitive coefficient matrix $\underline{K}$ represent respectively the inductive and the capacitive effects of the MTL. The current $|I>$ is equal to $(c|Q_L>)$ for z-directed propagation, c = speed of light, which gives the corresponding charge definition

$$|Q_L> = \underline{K}|V> \tag{5}$$

Thus $\underline{K}$ may be found from a two-dimensional capacitive boundary value problem. The matrix $\underline{L}$ is then given by the inverse

$$\underline{L} = \frac{1}{c^2}\underline{K}^{-1} . \tag{6}$$

The solutions to (3) and (4) may be written as

$$|V> = \sum_{i=1}^{N} (a_i e^{-j\beta z} + b_i e^{j\beta z}) \phi_i > \tag{7}$$

and

$$|I> = \sum_{i=1}^{N} (a_i e^{-j\beta z} - b_i e^{j\beta z}) c\underline{K}|\phi_i > \tag{8}$$

where $\beta = \omega/c$ and the summation is over the N voltage modes $|\phi_i>$. The $|\phi_i>$ are orthogonal and are normalized in the following inner product sense for unit power:

$$c < \phi_i | \underline{K} | \phi_j > = \delta_{ij}. \tag{9}$$

It is convenient to choose the $|\phi_i >$ to be the eigenvectors of $\underline{K}$ to give

$$\lambda_j c < \phi_i | \phi_j > = \delta_{ij} \tag{10}$$

where λ_j is the j^{th} eigenvalue of $\underline{K}$. With these definitions, the net power at any point on the MTL is given by

$$P = \tfrac{1}{2} \sum_{i=1}^{N} (|a_i|^2 - |b_i|^2). \tag{11}$$

Eqs. (7) and (8) may be written in more compact form as

$$|V > = \underline{M}_V(|a > e^{-j\beta z} + |b > e^{j\beta z}) \tag{12}$$

and

$$|I > = c\underline{KM}_V(|a > e^{-j\beta z} - |b > e^{j\beta z}) \tag{13}$$

and modeled as in Fig. 3 where

$$\underline{M}_V = [|\phi_1 >, \ldots, |\phi_N >].$$

The matrices $\underline{\Phi}_L$ and $\underline{\Phi}_R$ of Fig. 3, represent the phase delay between the aperture and terminations, $\underline{\Gamma}_3$ and $\underline{\Gamma}_4$. The traveling wave sources $|a_s >$ and $|b_s >$ represent the aperture coupling to the MTL.

The sources are obtained using the reciprocity relation

$$\int_V (E_a \cdot \bar{J}_b - \bar{H}_a \cdot \bar{J}_{mb} - \bar{E}_b \cdot \bar{J}_a + \bar{H}_b \cdot \bar{J}_{ma}) dv$$

$$= - \int_{\partial V} (\bar{E}_a \times \bar{H}_b - \bar{E}_b \times \bar{H}_a) \cdot d\bar{s} \ . \tag{14}$$

where the m subscript refers to magnetic current sources. For each voltage mode $|\phi_i >$ the corresponding electromagnetic field intensities are $\bar{e}_i$ and h_i respectively, where for $\eta = \sqrt{\mu/\varepsilon}$

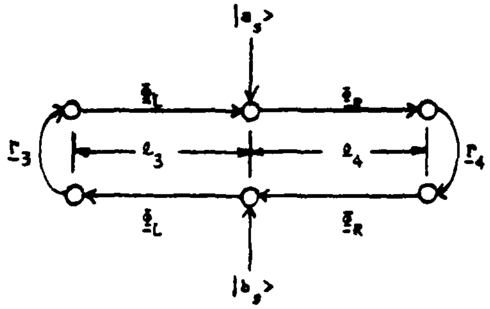

Figure 3. The zeroth-order signal flow graph.

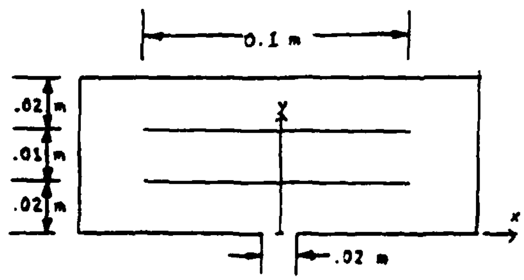

Figure 4. Cross section of two-conductor transmission line.

$$\eta \bar{h}_i = \hat{z} \times \bar{e}_i \ . \tag{15}$$

The total $\bar{E}$ and $\bar{H}$ have expansions similar to (7) and (8). Letting $\bar{E}_a$, $\bar{H}_a$, $\bar{J}_a$, and $\bar{J}_{ma}$ be equal to $\bar{E}$, $\bar{H}$, $\bar{c}_e \delta(\bar{r})$, $c_m \delta(r)$ respectively ($\delta(\bar{r})$ the 3-D Dirac delta), then using $\bar{e}_i$ and $\bar{h}_i$ with $e^{\pm j\beta z}$ for $\bar{E}_b$ and $\bar{H}_b$,

$$a_{s_i} = \tfrac{1}{2}[-c_{e_y} - c_{mx}/\eta] e_{y_i}(\bar{0}) \tag{16}$$

and

$$b_{s_i} = \tfrac{1}{2}[-c_{e_y} + c_{mx}/\eta] e_{y_i}(\bar{0}), \tag{17}$$

or equivalently

$$|a_s > = \tfrac{1}{2}[-c_{e_y} - c_{mx}/\eta] |e_y > \tag{18}$$

and

$$|b_s > = \tfrac{1}{2}[-c_{e_y} + c_{mx}/\eta] |e_y > \ . \tag{19}$$

Once the distances ℓ_3 and ℓ_4 and the termination voltage reflection matrices $\underline{\Gamma}_3$ and $\underline{\Gamma}_4$ are known, the total termination voltages may be obtained. These reflection matrices are given by

$$|b > e^{+j\beta\ell_4} = \underline{\Gamma}_4 |a > e^{-j\beta\ell_4} \tag{20a}$$

and

$$|a > e^{-j\beta\ell_3} = \underline{\Gamma}_3 |b > e^{+j\beta\ell_3} \tag{20b}$$

and the delay matrices by

$$\underline{\Phi}_R = e^{-j\beta\ell_4} \ \underline{I} \tag{21a}$$

and

$$\underline{\Phi}_L = e^{-j\beta\ell_4} \ \underline{I} \tag{21b}$$

where I is the N × N identity matrix. Concentrating on the termination at ℓ_4,

$$|V_4> = \underline{M}_V[\underline{I} + \underline{\Gamma}_4][\underline{I} - \underline{\Gamma}_3\underline{\Gamma}_4 e^{-j2\beta\ell}]^{-1}$$

$$\times \; [\,|a_s> + \underline{\Gamma}_3 e^{-j2\beta\ell_3}|b_s>\,]e^{-j\beta\ell_4}. \qquad (22)$$

$\underline{M}_V$ is the matrix of (12) and the second term accounts for the sum of $|a>$ and $|b>$. The inverse term accounts for the multiple reflections on the line with ℓ corresponding to the total line length. The remaining terms account for the source. The corresponding current may be obtained simply by replacing $\underline{M}_V[\underline{I} + \underline{\Gamma}_4]$ of (22) by $c\underline{KM}_V[\underline{I} - \underline{\Gamma}_4]$.

Bounds

The problem may now be stated as obtaining the upper bound of the voltage ΔV between any two wires at $z = \ell_4$. This voltage can not exceed twice the voltage of any wire to the ground plane. Thus

$$|\Delta V|_{max} \leq 2\left\| \, |V_4> \right\|_\infty$$

where the infinity norm of $|V_4>$ is the maximum absolute value $|V_{4_i}|$ over the wires. However, the infinity norm of a vector is less than or equal to the two norm given by

$$\left\| \, |V_4> \right\|_2 = \sqrt{|V_{4_1}|^2 + \ldots + |V_{4_N}|^2}$$

to give

$$|\Delta V|_{max} \leq 2\left\| \, |V_4> \right\|_2 \qquad (23)$$

To complete the bounding process, three important properties of matrix norms which follow are needed:

$$\| \underline{A} \| = 0 \text{ if and only if } \underline{A} = 0$$

$$\| \underline{A} + \underline{B} \| \leq \| \underline{A} \| + \| \underline{B} \|$$

$$\| \underline{A}\, \underline{B} \| \leq \| \underline{A} \| \; \| \underline{B} \| ,$$

along with the two norm of a matrix

$$\| \underline{A} \|_2 = [\max(\text{eigenvalue } \underline{A}^{\dagger}\underline{A}]^{\frac{1}{2}}$$

where $\underline{A}^{+}$ is the conjugate transpose of $\underline{A}$.

For passive terminations, the norms of both $\underline{\Gamma}_3$ and $\underline{\Gamma}_4$ are less than or equal to unity. With the norm of $\underline{M}_V$ given in terms of λ_i as

$$\| \underline{M}_V \|_2 = 1/\sqrt{c \min_i(\lambda_i)} \; ,$$

the upper bound on ΔV becomes

$$|\Delta V|_{max} \leq \frac{4}{\sqrt{c \min_i(\lambda_i)}} \; \frac{\| \, |a_s> \|_2 + \| \, |b_s> \|_2}{(1 - e^{-\sigma T})} \tag{24}$$

where σ_T represents the minimum round-trip loss on the MTL associated with the multiple reflections. Substituting for $|a_s>$ and $|b_s>$ from (18), (19), (1), and (2), this upper bound becomes

$$|\Delta V|_{max} \leq \frac{4\omega\varepsilon \, \| \, |e_y> \|_2 [|\alpha_e| \, |E_y^{sc-}| + |\alpha_{m,xx}| \, |\eta H_x^{sc-}|]}{(1 - e^{-\sigma T})\sqrt{c \min_i \lambda_i}} \tag{25}$$

The corresponding maximum current $|I|_{max}$ is simply given by

$$|I|_{max} = \frac{c|\Delta V|_{max}}{2} \; \sqrt{\max_i(\lambda_i)\min_i(\lambda_i)} \; . \tag{26}$$

Example

As a simple example of using (25), we will consider the dual strip-line geometry of Fig. 4 as described by Kajfez. The $\underline{K}$ for this geometry may easily be obtained from two-dimensional electrostatics [8] as

$$\underline{K} = \varepsilon_0 \begin{bmatrix} 15 & -10 \\ -10 & 15 \end{bmatrix}$$

with

$$\lambda_1 = 5\varepsilon_0 \ , \ \lambda_2 = 25\varepsilon_0 \ .$$

The appropriately normalized voltage eigenvectors of $\underline{K}$ are given by

$$|\phi_1> = \sqrt{\frac{\eta_0}{10}} \begin{bmatrix} 1 \\ 1 \end{bmatrix}, \quad |\phi_2> = \sqrt{\frac{\eta_0}{50}} \begin{bmatrix} 1 \\ -1 \end{bmatrix} .$$

The normalized electric field term is obtained by multiplying the $|\phi_i>$ by $\underline{K}$ to obtain the corresponding $|Q_{L_i}>$ from (5). Reverting to the electrostatic analysis, the short circuited field at the aperture may be obtained to give

$$|e_y> = -\frac{1}{.02} \begin{bmatrix} \phi_{12} \\ \phi_{22} \end{bmatrix} = \begin{bmatrix} -307.0 \\ 137.3 \end{bmatrix} .$$

The remaining parameters are chosen to be

$$\sigma_T = 90.652 \times 10^{-3} \text{ (or 1dB/100 ft.)}$$
$$E_y^{sc-} = 100 \text{ kV/m}$$
$$\eta H_x^{sc-} = E_y^{sc-}$$

and

$$a = \text{aperture radius} = 1 \text{ cm}$$

where $\eta_0 = 120\pi$ and the line lengths ℓ_3 and ℓ_4 are 7 and 5 meters respectively as used by Kajfez [8]. A circular aperture of radius 1 cm results in polarizabilities of $\alpha_e = 6.67 \times 10^{-7}$ and $\alpha_{m,xx} = 2\alpha_e$ [7]. The short circuited fields, E_y^{sc-} and H_x^{sc-}, correspond to a plane wave traveling along the surface.

Substituting in (25), the termination voltage for a radian frequency ω will not exceed

$$|\Delta V|_{max} \leq (2.38 \times 10^{-7})\omega \text{ V}.$$

This forms a useful frequency-domain bound for the problem presented with a voltage less than 10 Volts for frequencies below 6.7 MHz.

For comparison, this problem has been solved exactly for open-circuit terminations on the MTL. For such a case

$$\underline{\Gamma}_3 = \underline{\Gamma}_4 = \underline{I} \ .$$

Determining the $|a_s\rangle$ and $|b_s\rangle$ of Eq. (22) from (1), (2), (18) and (19), the results were computed and are plotted in Fig. 4 along with the bound. A modified bound is also plotted which represents the actual bound of the particular problem. The difference in bounds is 4.1 which seems slightly unreasonable until the bounding approach is examined. A factor of two arises in the bound to account for a differential mode which does not occur in the case considered. The triangle inequality used in the bound of $|a_s\rangle$ and $|b_s\rangle$ contributes another 1.5. The product of the 2-norms of $\underline{M}_V$ and $|e_y\rangle$ versus the ∞-norm of $(\underline{M}_V|e_y\rangle)$ contributes a 1.29 factor. A small contribution also occurs due to some of the neglected loss terms. In light of these observations, the resultant bound is very reasonable.

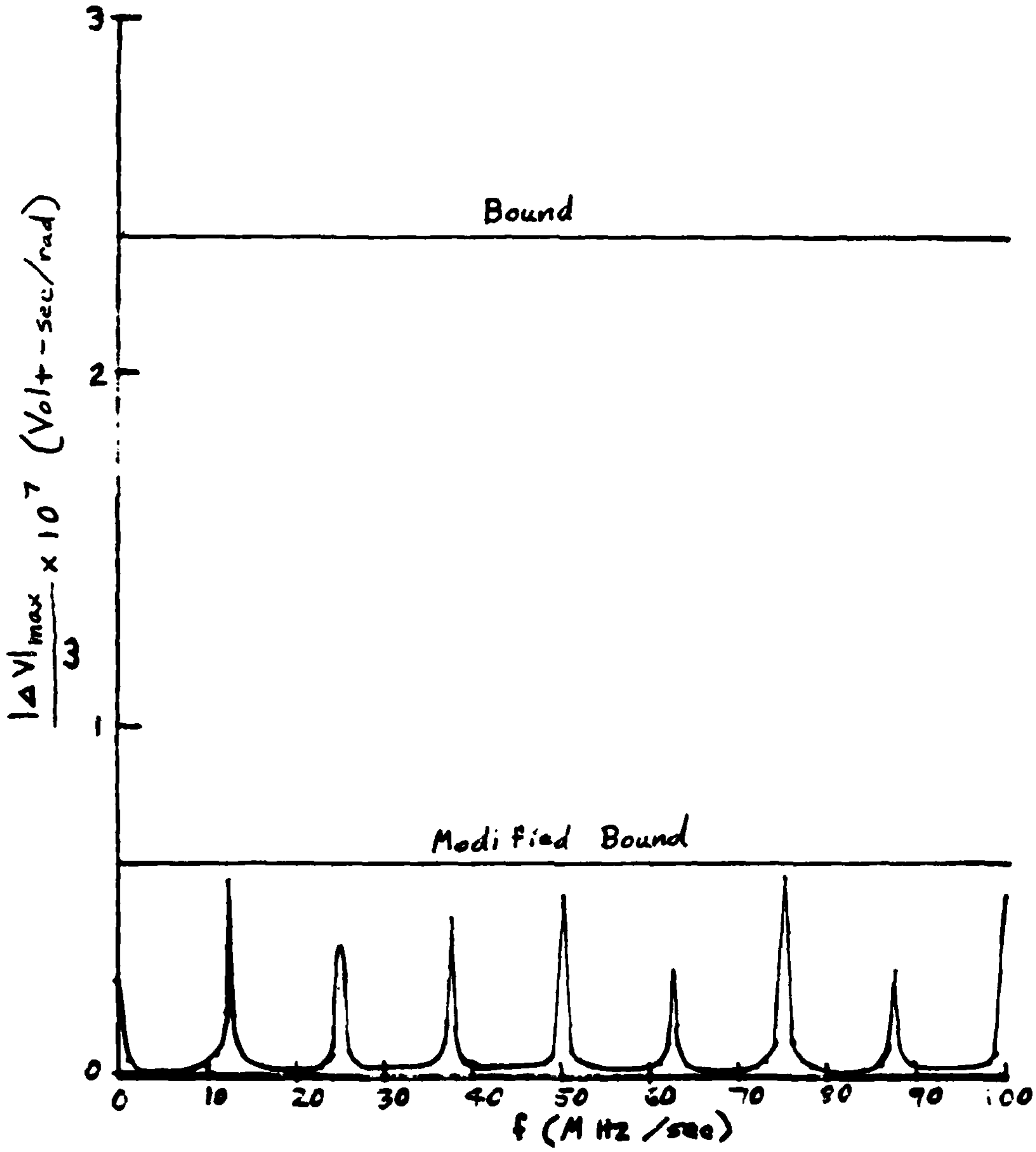

Figure 4. Open-circuit example.

Conclusion

This paper has presented an approach to bounding signal levels at MTL terminations due to excitation by an aperture. The approach requires knowledge of the external fields, maximum aperture dimension, cross section of the MTL, and the length of the MTL. The results use no knowledge of the terminations. Such an approach is useful for system hardness evaluations in aircraft and other systems to EMP or other incident energy.

Acknowledgement

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